SOLID STATE X-BAND SWEPT FREQUENCY TRANSMITTER INVESTIGATION

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THESIS

SOLID STATE X-BAND

SWEPT FREQUENCY TRANSMITTER

INVESTIGATION

by

Lt. Clifford J. Appel, USCG

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Solid State X-Band Transmitter

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INVESTIGATION

by

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ABSTRACT

The research conducted is an attempt to develop a swept frequency X-band transmitter for use as an aid to navigation. The investigation involves mounting an IMPATT diode in microstrip and stripline circuits to obtain a minimum rf output power of one watt. Provisions for electronic tuning and pulsing are also discussed.

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I. INTRODUCTION

The U. S. Coast Guard has been using a device as an aid to navigation known as a radar responder beacon, or racon. Its operation is analogous to IFF in that an interrogating radar signal causes the beacon to transmit a Morse coded pulse which can be uniquely identified on the radar PPI. The racon differs from ordinary beacons in that it transmits on a swept range of frequencies, one frequency of which will be that of the interrogating radar. The sweep rate must be slow enough to ensure that the transmitted pulse from the racon will fall within the passband of the interrogator's receiver during the antenna scan period. The Coast Guard currently uses the Vega Precision Laboratories Model 227-X racon, the specifications of which are listed in table 1.

Horizontally polarized radar signals are received by a broadband biconical horn antenna, go through a circulator, and encounter an interdigital filter which passes any rf signal from 9.3 to 9.5 GHz. A limiter prevents high level signals from damaging the low noise detector. A wideband video amplifier, which is always energized by a subordinate power supply, receives the pulses from the detector. After integrating about 40 pulses, a voltage threshold is reached which causes a turn-on circuit to energize the main power supply. The main supply feeds the Morse coder unit, sweep circuit and modulators, and transmitter. The main supply remains on until about 40 seconds after the last pulse from the interrogating radar has been received. While energized the transmitter responds with a Morse coded pulse for each radar pulse received, and at the same time, its frequency is slowly swept across the 9.3 to 9.5 GHz band. The output is fed through the circulator to the antenna. During transmission and for about 100 μsec thereafter, the trigger out of the video amplifier is inhibited in order to prevent

self-triggering. The transmitter output pulse is modulated in the form of one of the Morse characters D, G, K, M, N, or O. The theoretical maximum useable range of the racon is about 20 nautical miles, dependent upon racon height above sea level and the height of the interrogating radar's antenna. Since the delay time between a radar pulse and a racon response is short, the racon can supply a range, bearing, and positive target identification [1].

The racon described has been in operation for about four years. Some transmitters have failed during this time but have not been repaired. The transmitter consists of an L-band transistor energy source, for which no replacement is available, and the X-band signal is obtained through a varactor multiplier. It is the desire of the Coast Guard to update the racon transmitter with a solid state, X-band rf source, capable of an output power of about one to three watts.

The purpose of this investigation is to determine the feasibility of a replacement transmitter using an IMPATT diode in a microstrip or stripline configuration.

TRANSMITTER:	
Frequency	9310 to 9490 MHz $\pm$ 10 MHz
Frequency Sweep Interval	Selectable: 10, 30, 60, 90, or 120 secs.
Output Power	250 mw, minimum
Pulse Length	Adjustable: 25 to 50 μsec
Pulse Rise Time	0.05 μsec, maximum
Internal Delay	0.5 μsec, maximum
RECEIVER:	
Bandwidth	9.3 to 9.5 GHz
Sensitivity	-40dBm for 100% response
Radar Pulse Width	0.2 to 1.0 μsec
ANTENNA:	
Vertical Beamwidth	30° $\pm$ 5°
Horizontal Beamwidth	Omniazimuthal, within $\pm$ 2 dB
Gain	5 dB
POWER CONSIDERATIONS:	
Power Supply	12 V $\pm$ 2 V
Power Consumption	300 mw, minimum 1.5 w, maximum
OTHER:	
Environment Temperature	-40°C to +70°C
Weight	10 lbs., approx.
Height	16 inches including 3 inch spike
O.D.	8 1/2 inches
MTBF	30,000 hours, estimated

Racon Specifications

Table 1

II. TRANSMITTER REQUIREMENTS

Since the conception of the Model 227-X racon, microwave solid state power generators such as IMPATT, TRAPATT, and Gunn diodes have been developed which can produce higher outputs directly at the microwave frequency desired rather than multiplying the output from a lower frequency transistor. With the possibility of using one of these devices in a new transmitter, the Coast Guard established specifications for a new unit as shown in table 2. The new transmitter would function similarly to the old one by slowly sweeping from 9.3 to 9.5 GHz, eventually radiating on a frequency capable of reception through the receiver passband of the interrogating radar. Note, however, that it is possible for the sweep circuit in the transmitter to pass through the radar receiver's passband frequencies without the racon transmitting a pulse, such as in the case when the radar's antenna is not looking at the racon, therefore not providing the latter with a triggering pulse. Slowing the sweep rate might decrease the probability of such an occurrence but then the information rate becomes too low to be of any value for navigation.

The ideal solution would be to have a transmitter that radiates only on the frequency of the received pulse from the interrogating radar. Such a repeater beacon could be made in one of two ways. One method would be to determine the frequency of the interrogating radar with an instantaneous frequency discriminator[2,3] and use the output of the discriminator to set the racon transmitter frequency. This technique would do away with the slow sweep circuit but would still require the code modulator and pulsing circuits used in the swept frequency transmitter. The second method would be to use a reflection amplifier in a loop similar to the loop TWT used in ECM circuits[4]. Using this technique, the pulse received from the radar is

stretched, amplified, and re-radiated by the reflection amplifier and loop circuits in the racon. The stretched pulse could be code modulated during transmission. The sweep and pulsing circuits in the original racon would be unnecessary.

Neither of the two methods were pursued in a laboratory investigation. Instead, a new transmitter for the racon was attempted based on the specifications given in table 2.

Electronically swept (voltage or current) 9.3 to 9.5 GHz
Pulse width: adjustable (settable) 10 to 50 μsec.
Pulse rise and fall times: 50 nsec
Efficiency (DC to RF): 5%
Delay between modulating wave and RF envelope: 50 nsec.
Sweep rate: adjustable (settable) 200 MHz/sec to 2000 MHz/sec
Power output: 1 watt
Power stability: 0 to +10% for any part of sweep over temp. range
PRF: 1000 pps
Temperature: -20°C to +70°C
Freq. stability: 9305 to 9495 MHz $\pm$5 MHz over temp range
Other: PC board stripline modular construction will be required in production. Will be contained in a hermetically sealed enclosure with only ambient cooling. Production model will be shock and vibration tested.

Transmitter Specifications

Table 2

III. TRANSMITTER DESIGN APPROACH

A. SELECTION OF MICROWAVE POWER DEVICE

The choice of a power generating device was limited to using either a Gunn diode or an IMPATT diode. Industry has demonstrated that use of these devices is more straightforward than that of the relatively experimental TRAPATT diode for which the circuitry is somewhat more tricky to design. At the time of this investigation, however, Gunn diode technology had not yet produced a device which would meet the one watt minimum power output required at the duty cycle cited by the Coast Guard's specifications. Three manufacturers had IMPATT diodes which would meet these requirements. Also, a Gunn diode would have difficulty meeting the efficiency specification whereas a comparable IMPATT is about twice as efficient. For these reasons the Hewlett-Packard 5082-0425 silicon IMPATT diode was selected for this investigation. Table 3 lists the manufacturer's specifications.

Recently, however, Hewlett-Packard has made available a series of double-drift IMPATT diodes which have been shown to be more suitable for pulsed applications[5,6]. The double-drift diode is a P+PNN+ structure vice the P+NN+ structure of a single-drift IMPATT, thus enhancing the power output and efficiency. Since several of the 0425 diodes had been procured before existence of the double-drift diode and its advantages were known, the latter device was not obtained for experimentation.

V_{op}	100 V
I_{op}	210 mA
V_{BR}	78 V
C_j	.9 pF
C_p	.5 pF
L_p	.7 nH
η	7 %
Θ_{J-C}	7.0 °C/Watt

Manufacturer's Parameters
for 5082-0425 IMPATT Diode
Table 3

B. HEAT SINK DESIGN

The heat sink for the diode was designed to not only dissipate the heat produced but to also act as a mechanical support for the microwave circuit. The heat sink size can be determined by evaluating the thermal resistance necessary to transfer heat at the diode junction through the various mediums to the atmosphere as given by

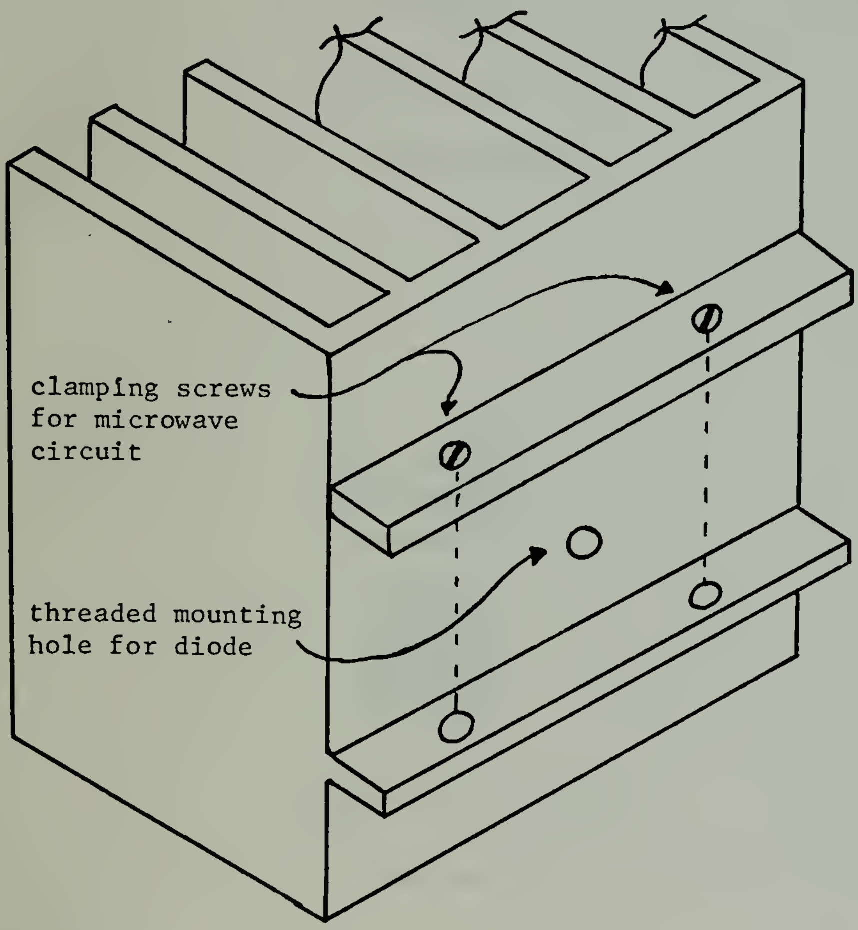
$$T_J = T_A + P(\Theta_{J-C} + \Theta_{C-HS} + \Theta_{HS-A})$$

where T_J is the diode junction temperature, T_A is the ambient temperature, P is the power to be dissipated by the diode, Θ_{J-C} is the thermal resistance between the diode junction and diode case, Θ_{C-HS} is the thermal resistance between the diode case and the heat sink, and Θ_{HS-A} is the thermal resistance between the heat sink and ambient air. Using the manufacturer's specifications, T_J =200 °C, P=21.0 watts, and Θ_{J-C} =7.0 °C/watt. T_A is nominally 25 °C and Θ_{C-HS} is nominally 0.2 °C/watt. Thus Θ_{HS-A} is calculated to be about 1.14 °C/watt when the diode is operated under normal cw conditions. During operation in the pulsed mode at a duty cycle of 5% as specified by the Coast Guard, the thermal resistance can be much higher allowing the physical size of the heat sink to be considerably smaller.

A heat sink was milled from a 3"x3"x1-3/4" block of aluminum by the metal shop of the Public Works Division at NPGS. A sketch of the sink is shown in figure 1. The thermal resistance is about 2.5 °C/watt, which can be reduced to safely handle the diode heat under cw conditions by forcing air at about 500 cubic feet/minute on the fins[7]. The forced air is not necessary when the diode is pulsed.

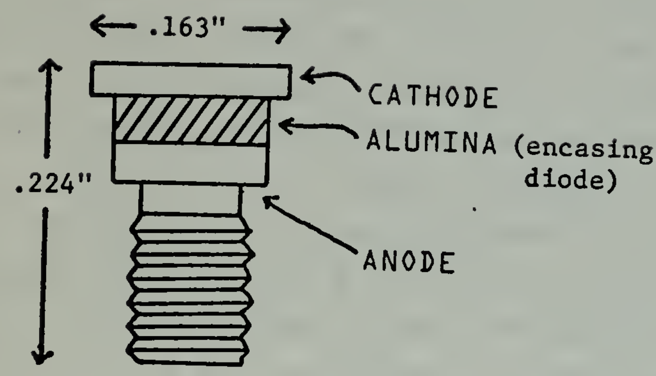
Figure 2(a) is a profile of the 0425 diode package. The anode end is threaded and fits snuggly into the hole in

the heat sink indicated in figure 1. The microwave circuit is then held in contact with the cathode end of the diode by screwing the clamping bar on top, squeezing the circuit together. This method held the circuit very securely. Figure 2(b) shows the chip mounted on the stud and connected to the cathode cap by a thin gold wire. The chip actually contains four diodes in parallel.



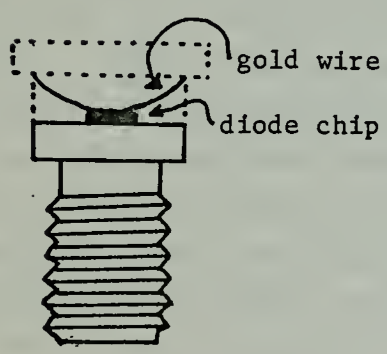
Heat Sink

Figure 1



Diode Package

Figure 2(a)



Diode Chip on Stud

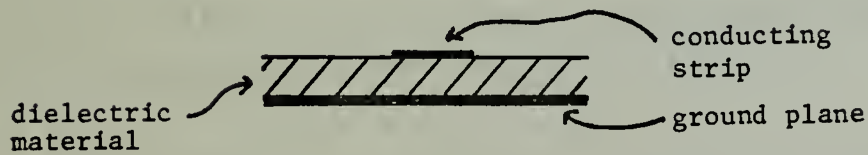
Figure 2(b)

C. STRIPLINE CONSTRUCTION

The term "stripline" is sometimes used in the general sense to describe any form of circuit construction consisting of a thin, flat conductor bonded to a dielectric substrate. For this investigation and for interpretation of the specifications given by the Coast Guard, stripline construction was construed as the triplate form shown in figure 3. This method consists of a circuit etched on a dielectric surface with another piece of the dielectric placed on top. On both the top and bottom of the dielectric a ground plane is formed by the unetched conductive material. This triplate construction is usually considered "stripline" and exhibits propagation of waves in the TEM mode. Another form of construction is microstrip, which is similar to stripline except that the top piece of dielectric and the top ground plane are non-existent, as shown in figure 4. Wave propagation in microstrip is in the hybrid mode due to the difference in permittivities above and below the circuit conductor. Many design graphs and nomograms have been generated for triplate stripline through computer programs and their accuracy has been documented by a large number of investigators using realized circuits verifying the theoretical calculations against measured data. Microstrip, on the other hand, is far more complicated to design due to the hybrid mode propagation, leaving some of the theory yet to be substantiated with measured data on actual circuits. For this reason stripline construction was exploited most of the time during this investigation.

Triplate Construction

Figure 3



Microstrip Construction

Figure 4

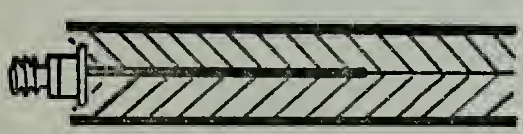
D. CIRCUIT BOARD AND MOUNTING

Two brands of printed circuit board were used during this investigation. The most used was Rexolite 1422, ϵ_r =2.55, .125 inch thickness. The Rexolite dielectric is a thermoset plastic covered on both sides with two ounce copper. When the circuit is sandwiched together in stripline form, the distance between ground planes is .250 inch, thus accomodating the cathode end of the diode quite well as it butts against the center conductor as shown in figure 5. Using the Rexolite the cathode fits between the ground planes since the cathode diameter is about .163 inch. The second type of board used was 3M CuClad teflon coated fiberglass, ϵ_r=2.50, .062 inch thickness. The dielectric is a woven glass cloth impregnated with teflon covered on both sides with one ounce copper. Thus, when this material is sandwiched, the thickness is .125 inch and part of the ground plane must be cut away to mount the diode using the procedure shown in figure 5 and to keep the cathode from shorting to the ground plane. The Rexolite was preferred over the 3M board since it was easy to machine. The glass from the 3M board tended to fray when the board was drilled and machined and these frays interfered with the construction process and electrical connections.

Two other methods of mounting the diode are shown in figures 6 and 7. The first method was used with stripline and the second was used with microstrip. The method in figure 6 was used with both the Rexolite and 3M board but the method in figure 7 for microstrip could only be used with the 3M material since the anode end of the diode would not fully protrude through the Rexolite to adequately screw into the heat sink.

The mounting procedure indicated in figure 5 is the preferred method for use with stripline circuitry since this

is analogous to mounting in a coaxial circuit[8]. The alternate procedure shown in figure 6 causes an unsymmetrical field pattern around the diode and is therefore not a desirable mount for a diode using this style of packaging[9]. This method was tried during the investigation to determine its effectiveness. The mounting style is, however, suited to microstrip circuits as demonstrated by other investigators[10,11].

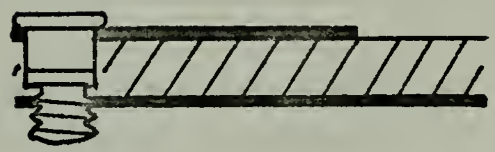


Diode End Mounting

Figure 5

Alternate Diode Mounting

Figure 6



Microstrip Diode Mounting

Figure 7

E. ORIGINAL OSCILLATOR DESIGN

Before designing the transmitter complete with a tuning device for sweeping 9.3 to 9.5 GHz, it was desired to first build an oscillator at 9.4 GHz with no provision for tuning. The purpose for the oscillator was to give the investigator a feel for using stripline design techniques and to check the operating characteristics of the diode.

The diode impedance was calculated using the manufacturer's parameters. A model of the IMPATT diode is shown in figure 8[12]. R_d is the negative resistance of the diode when reverse biased, C_j is the junction capacitance of the diode, L_p is the package inductance due to the gold wire connecting the diode to the cathode cap, and C_p is the package capacitance caused by the separation of the cathode cap and anode stud by the alumina encasement.

The manufacturer's curve of diode rf current as a function of R_d is shown in figure 9. From this curve calculation of power output as a function of R_d was made and plotted as indicated in figure 10. Thus, one watt output power or more can be attained at values of R_d between -1.4 and -.8 ohms. Using these parameters, the diode impedance between 9.3 and 9.5 GHz for values of R_d at -1.4, -1.2, -1.0, and -.8 ohms was calculated using the LISA program in the IBM 360 library. These impedances are plotted in figure 11 using the Smith chart as a negative resistance chart. A value of R_d=-1.2 ohms was chosen for the design since about 1.4 watts could theoretically be obtained according to the curve of figure 10.

A negative resistance device can function as an oscillator or an amplifier depending upon the impedance of the circuit in which the diode operates. If the resistive component of the circuit impedance is greater than the maximum absolute value of R_d, the device functions as an amplifier. If the real part of the circuit impedance is equal to $|R_d|$, oscillation will occur. The frequency of operation of either an amplifier or oscillator is determined when the circuit reactance is the conjugate of the diode reactance. With this in mind the normalized circuit impedance, $Z_{circuit}$, at 9.4 GHz was calculated to be +0.212-j1.324. The point $Z_{circuit}$ is plotted in figure 11 using the Smith chart as a positive resistance chart.

The design approach taken was suggested by the investigator's thesis advisor[13]. The thought was to use parallel coupled lines in order to keep the circuit length small and to keep the load isolated from the dc bias supply for the IMPATT. The load was chosen to be 50 ohms since this is the antenna impedance of the racon. It is not immediately obvious, but realistic arrangements of parallel coupled lines can be designed to transform 50 ohms down to values of between one and ten ohms only. An input to the parallel coupled lines of greater than ten ohms results in a gap spacing that is extremely close. The accuracy necessary to accomplish this spacing with the facilities at NPGS can not be easily achieved. On the other hand, an input of less than one ohm results in a wide spacing that results in very loose coupling. The gap spacing not only determines the input impedance, but it also determines the degree of coupling and hence the bandwidth of the parallel coupled lines[14]. A value of two ohms was arbitrarily selected as

the impedance to be used at the input to the parallel coupled lines.

Referring to figure 11 and starting at $Z_{circuit}$, rotation counterclockwise to the real axis results in ℓ =.102λ. The impedance is now pure real at about z_1 =13, Z_1 =650 ohms. A quarter wave transmission line can be used to transform the large resistance down to two ohms as the input to the parallel coupled lines.

$$Z_{MATCH} = (Z_1 Z_{IN})^{1/2} = (650 \times 2)^{1/2} = 36.1\,\Omega$$

The odd and even mode impedances (Z_{oo} and Z_{oe}) then need to be calculated to determine the proper strip width and gap spacing of the parallel coupled lines to transform two ohms to 50 ohms to match the load.

The equivalent circuit for the parallel coupled lines is shown in figure 12[15]. Since Z_{in} is real and θ is a quarter wavelength, then X_S =0 and Z_{in}=R_S =2 ohms. Thus

$$N^2 = \frac{Z_L}{R_S} = \frac{50}{2} = 25 \qquad N = 5$$

$$Z_{oe} - Z_{oo} = \frac{2Z_L}{N} = \frac{2 \times 50}{5} = 20$$

$$Z_{oe} + Z_{oo} = 2Z_L = 2 \times 50 = 100$$

Solving for Z_{oe} and Z_{oo}

$$Z_{oe} = 60 \qquad Z_{oo} = 40$$

Next the stripwidths for all the lines and the gap

spacing for the parallel coupled lines need to be determined. The values of w/b, strip width to ground plane spacing, and s/b, gap width to ground plane spacing, can be found by using the nomograms given by Cohn[16]. The line matching $Z_{circuit}$ to Z is a 50 ohm transmission line. Thus w/b is found for this single line by $Z_{oe}=Z_{oo}=50$ ohms. Rexolite 1422 was used for the initial design. Hence, $\epsilon_r=2.55$ and b=.250 inch. For the 50 ohm stripwidth, w/b=.75 and w=.188 inch. Similarly, the quarter wave matching line of 36.1 ohms has a w/b=1.24 and w=.310 inch. Using Z_{oe} and Z_{oo} calculated for the parallel coupled lines, w/b=.7 and s/b=.135. Thus, w=.175 inch and s=.0337 inch.

The last step is to determine the lengths of each of these lines. The wavelength in the dielectric at 9.4 GHz is

$$\lambda' = \frac{\lambda_o}{\sqrt{\epsilon_r}} = \frac{3\times10^8 \text{ m/sec}}{(9.4\times10^9)(\sqrt{2.55})} = 3.19 \text{ cm} = 1.255''$$

The length of the 50 ohm line, .102 λ, is then .081 inch. The quarter wavelengths are .198 inch.

Provision must be made to bias the IMPATT. This is done through a bias tee which does not affect the impedance of the surrounding circuitry. The point at which the bias line touches the IMPATT and its circuit should be a low impedance point. The bias line should be of high impedance (in parallel at the point of contact) so as to not alter the impedance at the point of contact. Thus, the bias line touching the diode circuitry is very thin, about ten mils wide, resulting in a high impedance. Then at a quarter wavelength away, the bias line is terminated by a much wider line, about .400 inch wide and a quarter wavelength long, resulting in a very low impedance.

A sketch of the circuit and biasing arrangement is shown in figure 13. It shows the relative sizes of each circuit element, but it is not drawn to perfect scale. The bias line was made 3/4 wavelength long in order to be routed conveniently away from the rest of the circuit. The line connecting the 36.1 ohm matching line to the parallel coupled lines was made 1/2 wavelength long in order to keep the bottom coupled line from touching the 36.1 ohm strip. The bottom line continues to the right to the edge of the board where it contacts the center tab of an SMA coaxial connector.

The pattern was laid out on a transparent sheet of plastic, 12 times the desired circuit size. This method not only insures precision dimensions when reduced, but it also enables the pattern to be constructed with more ease. The negative, reduced to actual working size, was then used to expose the photo-resist sprayed and baked on the Rexolite. The photo and developing process used was the exact process used for other types of printed circuit boards. The only departure was during the etching stage. It was observed that ferric chloride etchant produced a final circuit with far cleaner, less pitted lines than did the ammonium persulfate etchant commonly used on printed circuit boards. The SMA connector was attached to the output strip and held the right end of the Rexolite pieces together. Since this is a stripline circuit the bottom side of the board in figure 13 had all of its copper retained. Another piece of Rexolite placed on top of the circuit contained a ground plane on the outside and a fully etched surface on the opposite side so that the side view of the completed circuit appears as noted in figure 3. Before the two pieces were sandwiched a wire was attached to the point shown as B+ on figure 13. A hole was drilled through the mated pieces at the point indicated by the dotted lines so that the cathode of the diode would touch the circuit. Ideally, perfect

machining is necessary to get the diode to butt up against the circuit. However, good contact could not be made so a small tab was attached at the input strip and protruded a couple of mils assuring firm contact when the diode was positioned as noted in figure 5. This does in fact alter the electrical length of the input line and future circuits took the length of the tab into account during calculation. The diode and circuit were then placed in the heat sink which squeezed the left end of the pieces together and held the IMPATT and its circuit firmly together.

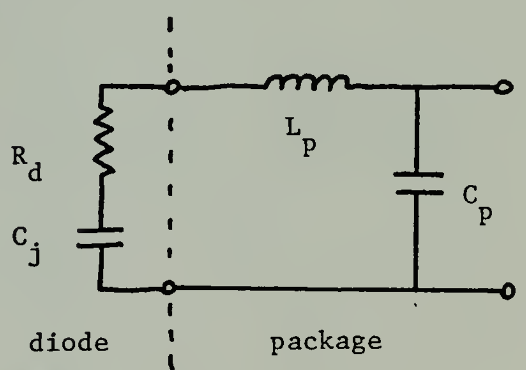
The circuit described did not enjoy much success. The output frequency displayed on an HP5340A digital frequency counter was about 6.237 GHz. The output power was about 59 milliwatts determined by an HP432A power meter. The frequency of oscillation for a diode dc current of 164 mA varied between 6.221 and 6.244 GHz depending upon the depth that the diode was screwed into the heat sink. The power also varied between 40 and 59 milliwatts. The differences in power and frequency are caused by the subtle change in diode impedance as a result of its position in the heat sink. The fact that the circuit performed at 6.237 GHz producing 59 mW instead of at 9.4 GHz at 1.4 W as calculated indicated that in the stripline environment, the diode parameters as given by the manufacturer are no longer applicable. It is also believed that the IMPATT could have been operating in the TRAPATT mode because of the poor circuitry. Another phenomenon encountered as the bias current was varied was that of jump and hysteresis[17]. At about 114 mA, the frequency jumped from 5.955 to 6.218 GHz and the power dropped from 18.8 to 4.8 mW. Then as the bias current was decreased, the jump occurred at about 109 mA. The frequency changed from 6.216 to 5.937 GHz and the power changed from 4.6 mW to 24.3 mW. The occurrence of jump and hystersis is also another indication of poor circuit design.

During the course of the calculations one important factor was neglected and omitted during circuit construction. The capacitive effects resulting from the abrupt discontinuities at the open end of each of the parallel coupled lines had been overlooked. Thus, the line appears electrically longer. This not only makes the top coupled strip in figure 12 longer than 3/4 wavelength, but also the coupled lengths of the top and bottom lines are greater than a quarter wavelength. The influence of end effects is a function of the width of the strip. Using the graph from previous investigations of end effects[18] for $w/b=.7$, then the end effect to ground plane spacing ratio, Δ/b, is about .158. (Some texts, such as Ref. 14, state that $\Delta/b=.165$ is a good rule of thumb for all strip widths.) Thus, $\Delta=.0395$ inch which means that each coupled strip was about 40 mils longer than it should have been.

The circuit was modified to the correct strip lengths by placing tape across all the copper lines except for leaving 40 mils exposed at the open circuited ends of the parallel coupled lines. The board was re-etched, removing the copper not covered by the tape. The circuit was once again assembled and tested. This time the frequency of oscillation was between 4.970 and 5.163 at 150 mA and the output power was between 100 and 140 mW, depending on the diode depth in the heat sink. The lower frequency and higher power outputs encountered were again thought to be that of the TRAPATT mode.

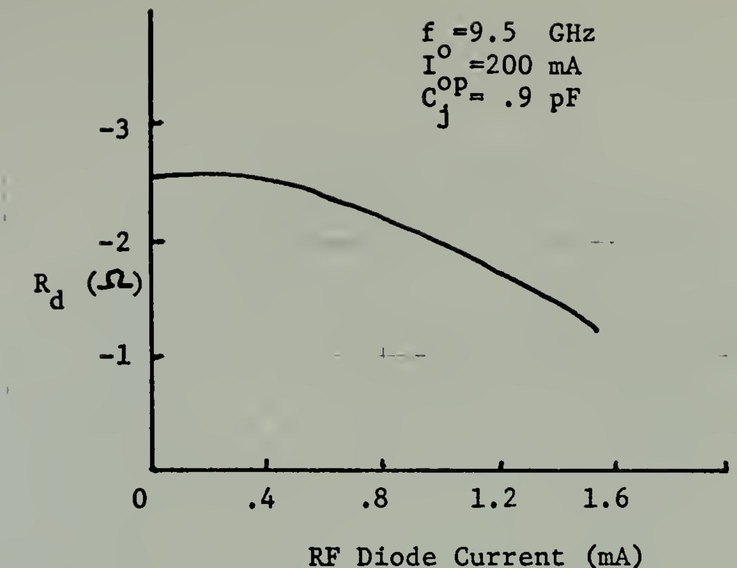
Even though the diode operating current is much higher than the 164 and 150 mA bias currents previously stated, further increase from these values led to a reduction of output power and frequency. This is due to the change of parameters R_d and C_j which became less than optimum for

matching to the circuit impedance. Also, the efficiency of the heat sink was being tested to insure the IMPATT's junction temperature was not exceeded. Bias currents around 190 to 200 mA were applied and the heat sink was found to be adequate as long as the 500 cfm fan was in operation. After several minutes of operation at these current levels, the sink was only warm to the touch. Apparently, it was the only device which had been designed correctly. As a point of interest, HP tests its 0425 diodes in a coaxial fixture attached to a 2"x2"x1" copper block which is welded to a larger aluminum plate 6"x8"x1/2"[9].



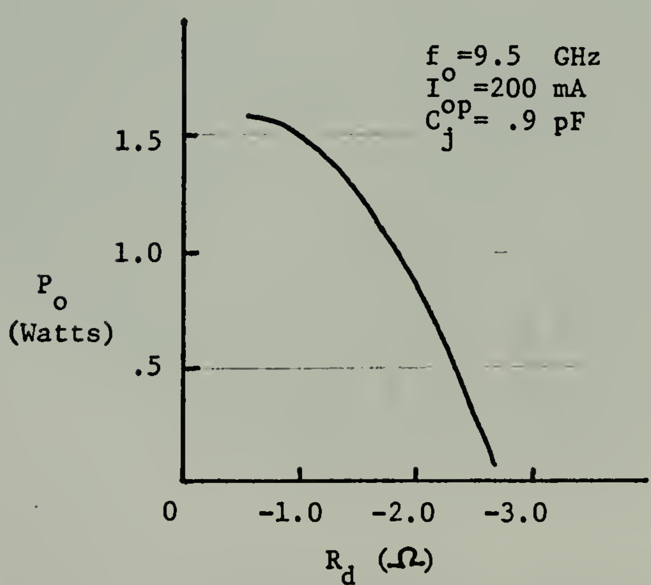
IMPATT Diode Model

Figure 8



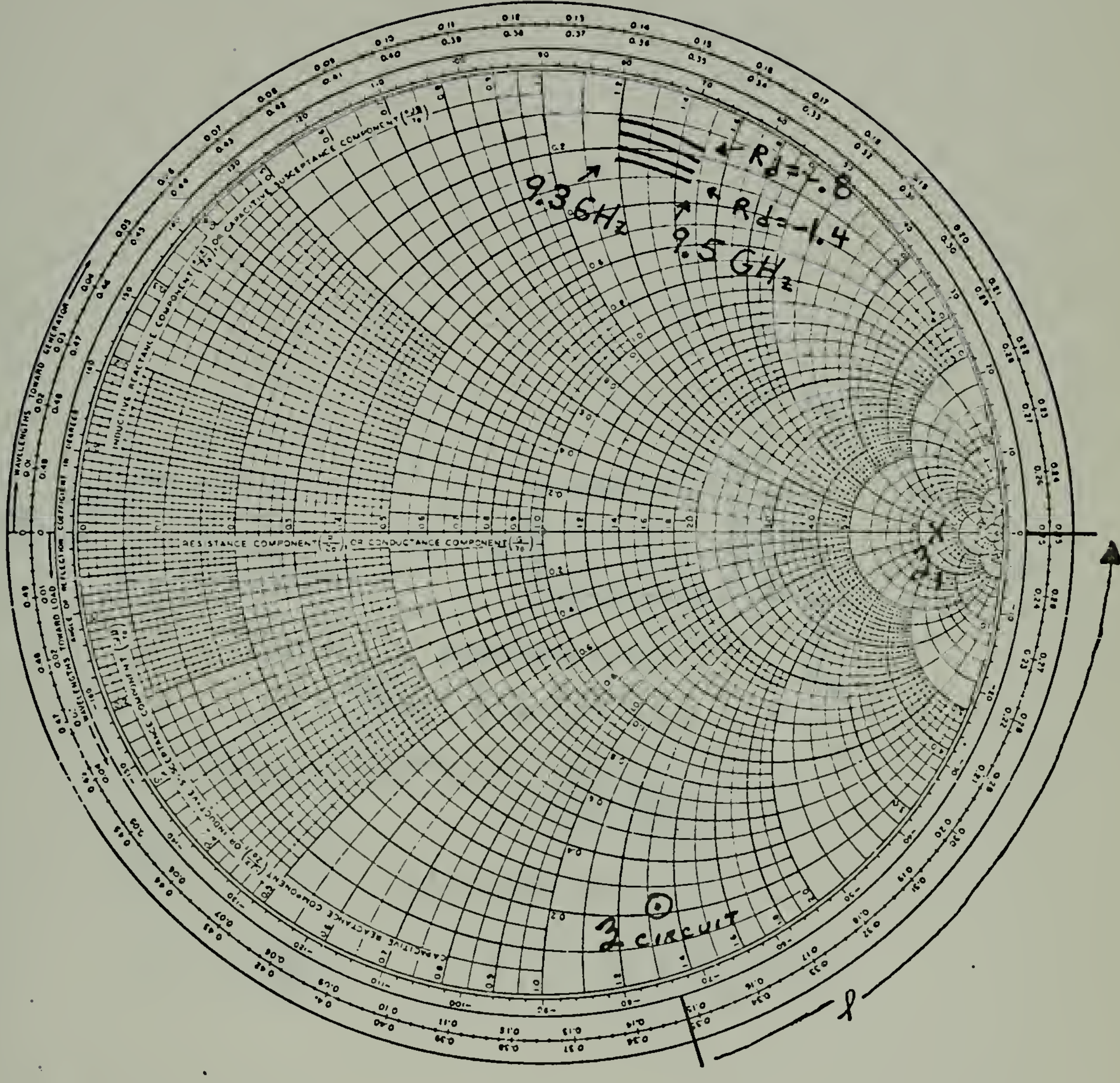
RF Current as a Function of R_d

Figure 9



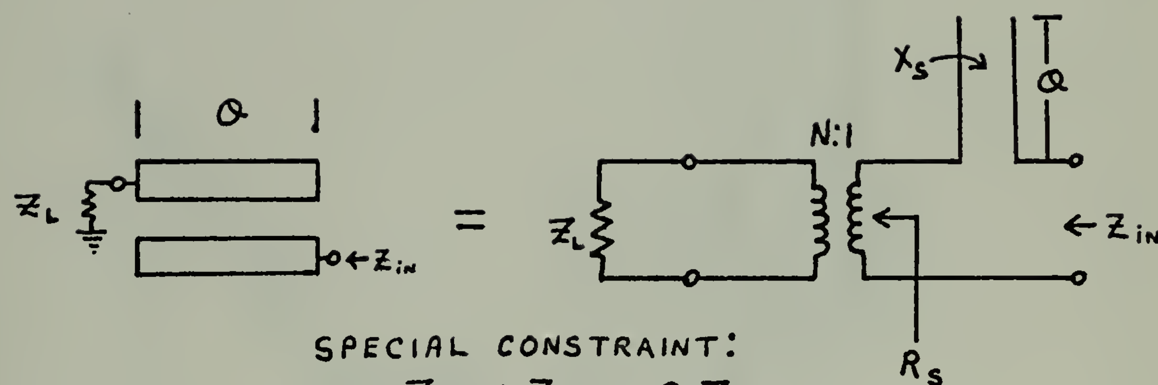
Power Output as a Function of R_d

Figure 10



Diode Impedances and Z_{in}

Figure 11



SPECIAL CONSTRAINT:

$$Z_{oe} + Z_{oo} = 2 Z_L$$

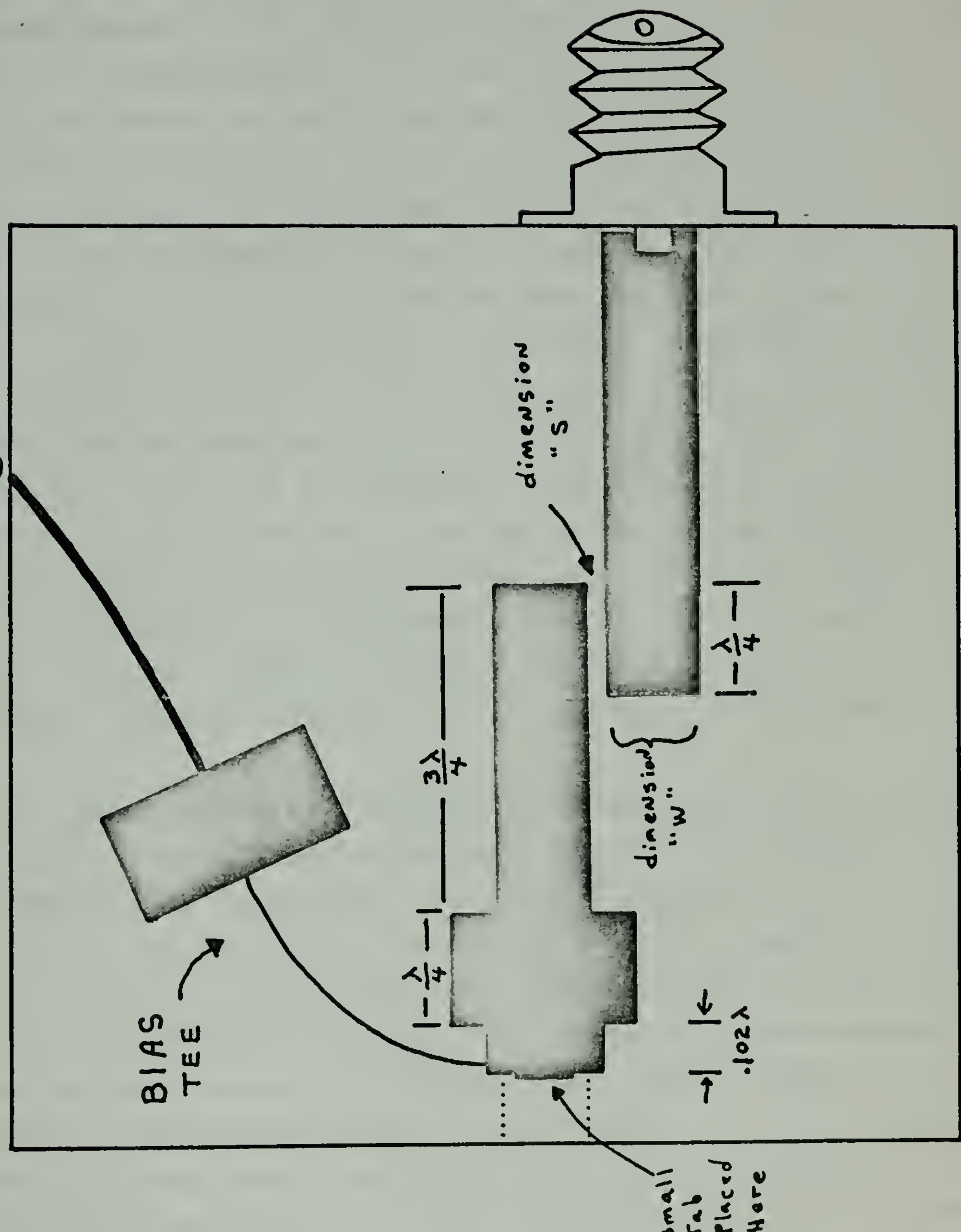
$$N = \text{TURNS RATIO} = \frac{2 Z_L}{Z_{oe} - Z_{oo}}$$

$$Z_{IN} = R_S - j X_S \cot \theta$$

$$R_S = \frac{Z_L}{N^2} \qquad X_S = Z_L \left[\frac{N^2 - 1}{N^2} \right]$$

Equivalent Circuit for Parallel Coupled Lines

Figure 12



Original Circuit

Figure 13

F. SUCCESSIVE DESIGN ATTEMPTS

The failure of the previous approach to design a 9.4 GHz oscillator with greater than one watt output power caused the investigator to resort to another method of design in order to determine the IMPATT's true impedance in a stripline environment. One reason for the failure of the first approach could be due to the fact that there are too many frequency dependent elements used in the design. It was suggested that the method used in Ref. 11 be tried in order to determine the diode impedance experimentally[19]. The procedure is to use end coupled lines, each having a characteristic impedance of 50 ohms. The line contacting the diode acts as a resonator, its length determining the frequency of oscillation. The gap between the resonator and the line connecting the load then determines the load on the IMPATT and couples the microwave energy to the load. This method is also acceptable since it keeps the dc bias out of the load. Figure 14 shows an end coupling arrangement.

A total of seven circuits were manufactured, each with varying resonator lengths and gap spacings determined by how successful the previous end coupled circuit had been. The frequencies of oscillation of each circuit varied between 3.900 GHz and 9.043 GHz. The largest output power was as high as 170 mW at 9.041 GHz. One peculiar phenomenon was noted during these attempts. One would expect that the frequency of oscillation would get higher as the resonator length is shortened, keeping the gap space constant. This, however, did not occur in two instances. These circuits were made on Rexolite and the diode was mounted as noted in figure 5.

Next, five more circuits were made using the 3M board and similar diode mounting procedure. Slots had to be made at the point at which the diode entered the board since the

diode is larger than the ground plane spacing. Not one of these circuits produced oscillations, even though the w/b and r/b (gap to ground plane spacing) ratios were almost the same as a few of the circuits using the Rexolite.

Four more circuits using Rexolite were made but in this case the diode was mounted as noted by figure 6. The dimensions for these circuits were the same as those of the previous Rexolite circuits. No oscillations were observed from any of the circuits.

It was thought that perhaps the diode packaging size might be so large that its reactance was overwhelming the relatively smaller dimensions of the strips. To test this hypothesis, a large square cavity .550 inch on a side was built on Rexolite to attempt to resonate the diode. Coupling from the cavity was made by a 1/2 wavelength line which was determined to be about 22 ohms, since w/b=.550/.250=2.2. Then a quarter wave matching line of 33 ohms was used to connect the line to a 50 ohm line which terminated at the load. Figure 15 is a sketch of this circuit. Four of these circuits were built, each with a different gap spacing between 20 and 60 mils. These spacings seemed to be successful for the first seven end coupled circuits on Rexolite. The diode was mounted in the perpendicular position. In each circuit, the IMPATT was first mounted at the edge of the resonator and then in the middle of the resonator. No oscillations were observed for any of the circuits.

At this point it was decided to determine the impedances for some of the circuits that did oscillate and also for those that did not work. The equivalent circuit for end coupled lines can be modeled in two ways. Figure 16(a) is the centerline representation and figure 16(b) is the edge representation[20]. Equations for calculating the

susceptances in each model are given in Ref. 20. In addition, the reference also gives graphs for the normalized susceptances based on the gap to strip width ratio, r/w. An example of an impedance calculation using the edge representation for one of the end coupled circuits follows.

A circuit that oscillated at 9.041 GHz at a power output of 170 mW had dimensions as follows: resonator length=.510 inch, r=.050 inch, and strip width=50 ohms=.188 inch. Then r/w=.266. Entering the curves given in Ref. 20

$$\frac{B_A}{Y_o} = .087\ \mho \qquad \frac{B_B}{Y_o} = .087\ \mho \quad \therefore \quad \frac{X_B}{Z_o} = 11.50\ \Omega$$

The calculation of the circuit input impedance from this point is no more than an undergraduate excercise in the use of the Smith chart. Thus, as noted in figure 17, starting at the center of the Smith chart, normalized load of 50 ohms, the first susceptance is added. The resulting admittance is converted to impedance and the series reactance, X_B/Z_o, is added. The resulting impedance is converted back to an admittance. The remaining shunt susceptance, B_A/Y_o, is added and the resultant admittance is converted back to impedance. This is the impedance at the gap end of the resonator. The circuit input impedance is found by rotating clockwise around the chart a distance equal to the resonator length in wavelengths. At 9.041 GHz, the wavelength in the dielectric is 2.07 cm = .818 inch. Thus

$$\ell = \frac{.510}{.818} = .624\lambda$$

The normalized circuit input impedance is about .02-j0.73, or $Z_{circuit}$ =1.0-j36.5 ohms. The impedance of the IMPATT in this particular circuit was therefore -1.0+j36.5 ohms. This value is somewhat different from the value calculated by using the manufacturer's parameters. It is interesting to note the low real part of the diode impedance with respect

to the larger imaginary part. This is also another indication that perhaps the package reactance was playing too large a role in determining the frequency of operation as theorized earlier.

Similar calculations were made for other circuits that oscillated and the values of circuit impedance for two of them are shown in figure 17. In a similar manner, the impedances for two circuits that did not oscillate are also shown. Their calculations were based upon the fact that they should have worked at 9.4 GHz. The large resistive values of both of these latter calculations indicates the reason that no oscillations were noted for the circuits. As with the parallel coupled lines, the larger the gap spacing, the lower the input resistance of the circuit, and vice versa. It was not known why the plots of circuit impedance at the indicated frequencies did not follow a logical pattern. This strange occurrence, along with the fact that shortening the resonator length did not always raise the frequency of operation, could be due to some unidentified parameters being introduced into the circuit by the presence of the diode.

Having been unsuccessful at obtaining 9.4 GHz using the end coupling technique, another plan of attack was formulated. Various diode impedances were assumed, and consequently circuits were designed to match the assumed impedances. The circuits were constructed using parallel coupled lines, the input line acting as a resonator with no change in strip width occurring down the length of the resonator. The circuits were designed using the equations shown in figure 12 and the length of the resonator determined the reactive part of the circuit input impedance. The second circuit designed based on an assumed impedance resulted in oscillations of 46 mW at 9.301 GHz and the third circuit based on an assumed impedance produced 105 mW at

9.558 GHz. The IMPATT impedance was concluded to be about -5.0-j19.5 ohms at 9.4 GHz when placed in a stripline environment of Rexolite.

The next exercise was then to determine the input resistance which would give the 1.4 watts anticipated. A total of 12 circuits were built in an attempt to pinpoint the exact value of impedance necessary. The largest output obtained was 130 mW from a circuit whose impedance was designed for 3.75+j21.0 ohms.

It was thought that the poor power response was due to the fact that the large ground plane spacing associated with Rexolite was attenuating the fields since the frequency of oscillation was higher than that normally required when using this material[8,21]. As a test of this hypothesis, the same type of circuit design was used in conjunction with the 3M board. In these instances, though, the strip widths and gap spacing are nearly half that of the ones using Rexolite. Thus, once again, the diode packaging is large relative to the strip widths. The gap spacing is far more critical also. Five circuits were built whose impedances were in the neighborhood of 3.75+j21.0 ohms but none of these yielded oscillations of any power level at any frequency.

One final attempt at an oscillator design was made in a microstrip configuration using the 3M board. The same value of circuit impedance mentioned previously was used as a starting point in a parallel coupled lines design. Thus, Z_{oe} and Z_{oo} were the same. The determination of strip widths, gap spacing, resonator length, and end effects for the even and odd mode impedances is quite different and far less precise for microstrip than for stripline since the

phase velocity of the hybrid wave changes with the stripwidth.

Starting with Z_{oe} and Z_{oo}, w/h and s/h (where h is the dielectric thickness) for a dielectric with ϵ_r=2.50 had to be found. Entering the quasi-static curves generated by Bryant and Weiss[22] for permittivities of 6.0, 9.0, and 12.0, horizontal lines at Z_{oe} and Z_{oo} were drawn across each graph. Thus, w/h and s/h were found at the points where s/h for each impedance was the same. These ratios were then plotted on semi-logarithmic graph paper as a function of ϵ_r. The values of w/h=2.2 and s/h=0.2 were extrapolated at the point ϵ_r=2.5. The resonator length was determined by noting the velocity for the calculated strip width as 2.15 X 10^8 m/sec[23]. The end effect for each line was determined by using the rule of thumb, Δ/h=.33[24]. The circuit was etched on 3M CuClad board, .062 inch thick.

The microstrip circuit failed to produce any oscillations. Three additional microstrip circuits were manufactured such that the real part of the circuit impedance was one ohm, five ohms, and seven ohms. None of these circuits oscillated either. It is interesting to note that none of the circuits built on the 3M board in either a microstrip or stripline configuration resulted in oscillations at any frequency.

It was suspected that the loss in the stripline circuit that produced 130 mW was quite large due to the wide ground plane spacing or poor rf coupling between the parallel coupled lines. If this hypothesis were true, then the following relationship would be true

$$|R|^2 + |T|^2 + |L|^2 = 1$$

where R is the reflection coefficient, T is the transmission coefficient and L is any loss that might be incurred.

The circuit was placed in the HP8743A Reflection-Transmission Test Unit. The transmission insertion loss was found to be -8dB on the HP8412A Phase-Magnitude Display and $|R|$ was found to be 0.9 on the HP8414A Polar Display. These units are all parts of the HP8410S Network Analyzer System driven by the HP8690B Sweep oscillator. The transmission coefficient, without loss, is then

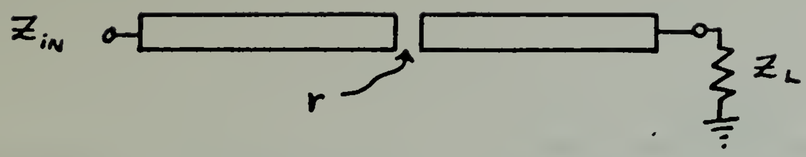
$$1 - |R|^2 = |T|^2 \Rightarrow |T|^2 = .19$$

The measured transmission coefficient, with loss, is

$$|T|^2 + |L|^2 = -8dB = .158$$

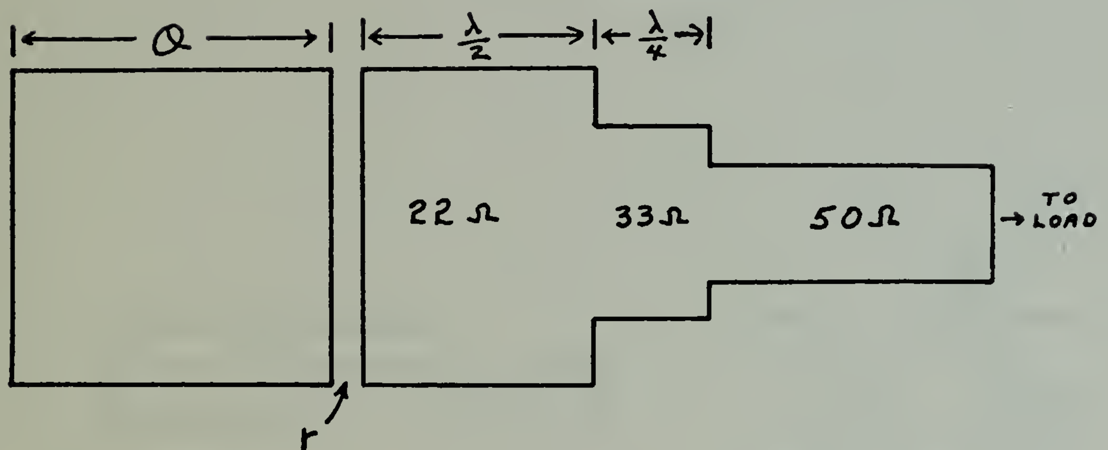
The power loss in the circuit can be concluded to be 0.032. Therefore, it is reasonable to assume that circuit losses are not responsible for the low power output of the parallel coupled lines circuit using Rexolite.

A suggestion was made to take advantage of the fact that the IMPATT can be made to resonate its own packaging[21]. The method would be to transform the 50 ohm load down to about one ohm through a seven ohm quarter wavelength transformer. This also puts the diode bias voltage on the load unless a biasing network, such as the HP11590A or similar arrangement, is used for dc isolation. But a seven ohm transmission line in stripline would be so much wider when compared to its length that the discontinuities at the line ends could not be accurately determined for proper circuit design. The technique is ideally suited for coaxial designs, however.



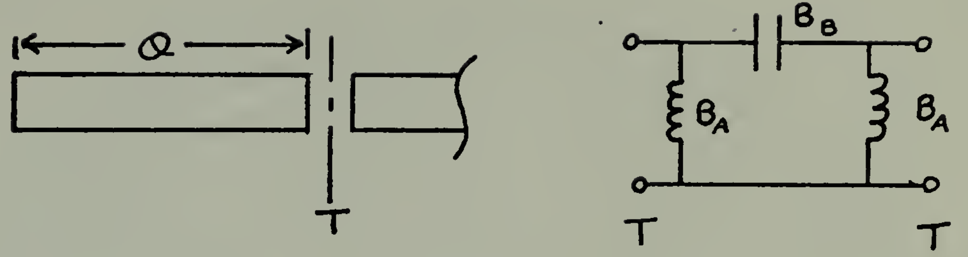
End Coupled Lines

Figure 14



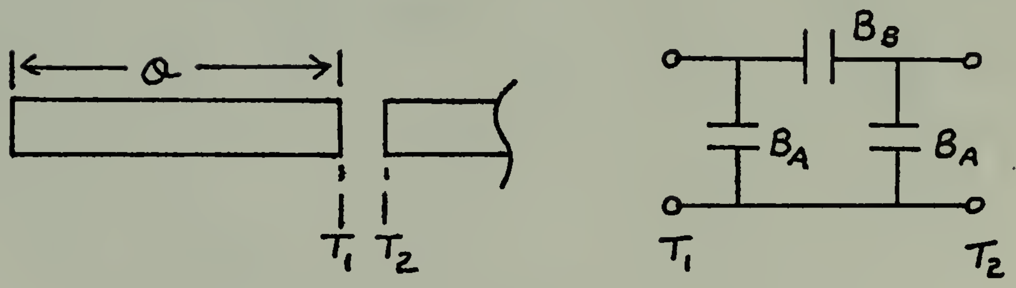
Square Cavity Circuit

Figure 15



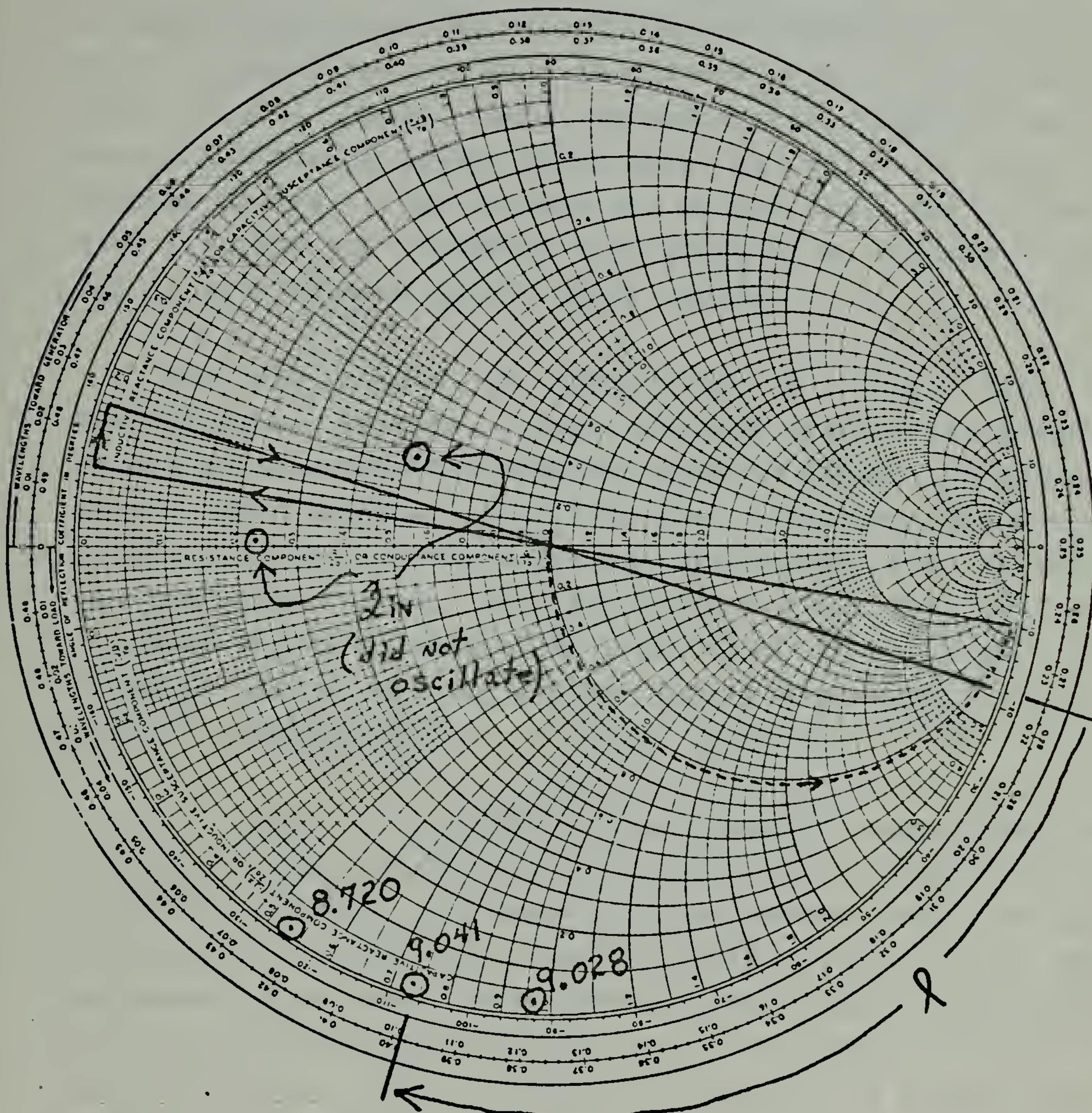
Centerline Representation

Figure 16(a)



Edge Representation

Figure 16(b)



End Coupled Impedance Calculation

Figure 17

G. TUNING DEVICE CONSIDERATIONS

Although the basic oscillator with more than one watt of output power was never realized, consideration for tuning such an cscillator had been researched and the groundwork laid. The Coast Guard specifications called for an electronic tuning device which must be able to tune the oscillator at least through 9.3 to 9.5 GHz. The choice of such a device was therefore made between a varactor diode and an yttrium iron garnet (YIG) sphere.

The YIG sphere is physically a tiny ball of ferromagnetic material whose internal crystalline field is varied by an externally applied magnetic field. These two fields add together to form the effective field which directly determines the resonance frequency. The varactor, on the other hand, is a PN junction diode that acts as a variable capacitor when a reverse bias is applied. The varactor is a voltage tuned device whereas the YIG is a current tuned device.

YIG tuned oscillators have several advantages over varactor tuned oscillators. Temperature stability of an oscillator using a YIG sphere is much better than one using a varactor. YIG spheres have a far wider tuning range than varactors and the former give a fairly linear tuning curve whereas the latter tunes frequency as the fourth root of the applied voltage. YIG tuned circuits also give higher cutput power due to their higher Q's and lower insertion loss. However, since a YIG tuned oscillator requires a magnetic field, the tuning response is slower than that of a varactor. The magnetic field needed for tuning requires dc power (and hence lowers the dc to rf conversion efficiency) whereas the varactor requires nearly no tuning power. Thus, a YIG circuit would be bulkier and heavier due to the magnetic-tuning inductor[25]. It was decided that the

IMPATT oscillator would be tuned by using a varactor diode.

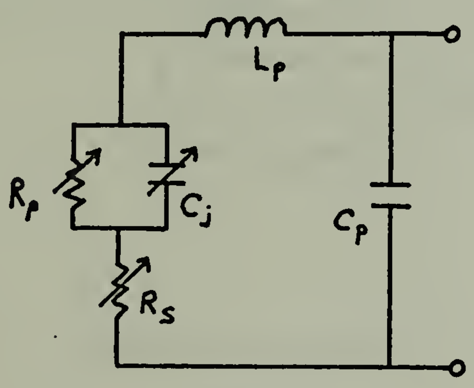
A model of a varactor diode and packaging is shown in figure 18[26]. C_j is the capacitance formed at the PN junction, R_p is the shunt resistance at the junction, usually considered negligible, R_s is the series resistance which accounts for loss in a varactor, and C_p and L_p are the package capacity and inductance respectively. The Q of a varactor is calculated by using parameters C_j and R_s. Thus, R_s is an important factor affecting a varactor's efficiency.

A varactor was procured in anticipation of a successful oscillator design. Referring to the Smith chart in figure 11, one sees that the reactance of the IMPATT varies about 14 ohms at R_d=-1.2 ohms between 9.3 to 9.5 GHz. A GHZ Devices GC-1500 varactor seemed to be more than adequate for tuning through this reactance range. The manufacturer specifies the measured Q to be about 3600 at 50 MHz, the capacitance at -4 volts reverse bias to be about .8 pF, and the capacitance tuning ratio to be about 3.3 over a bias voltage range of 30 volts. The reactance range of the diode is therefore between 21 and 70 ohms at 9.4 GHz. Since the range is greater than that actually needed, it was thought that linearity would be enhanced by using only a small portion of the varactor tuning curve.

The original design of the swept frequency oscillator was to employ three parallel coupled lines as shown in the schematic representation of figure 19. In order to keep the strip widths and gap spacings the same for all three lines to simplify design calculations, the varactor would be biased to provide 50 ohms reactance at the starting frequency of 9.5 GHz. The parallel combination of the 50 ohm reactance and the 50 ohm load resistance would form an

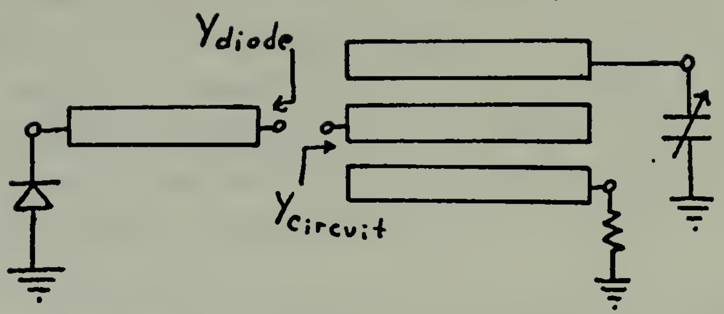
admittance that is the negative of the IMPATT admittance when transformed to a useable value for the IMPATT through the parallel coupled lines. The oscillator would tune from 9.5 to 9.3 GHz as the bias increased[13].

Had the approach using a resonator and parallel coupled lines been successful, the varactor could have been put in series with the IMPATT[10,11]. This technique involves placing the varactor at the low impedance point of the resonator such that the varactor changes the electrical length with biasing as shown in figure 20. The varactor bias would vary with respect to the IMPATT bias to accomplish tuning whereas in the parallel arrangement bias would vary with respect to ground. Each method has its own advantages. The series arrangement has been found to provide a greater tuning bandwidth[27] whereas a varactor arrangement in parallel with the IMPATT results in a greater output power[28]. However, the efficiency and output power of either arrangement can be expected to be somewhat less than a circuit without a varactor as found by other investigators[29].



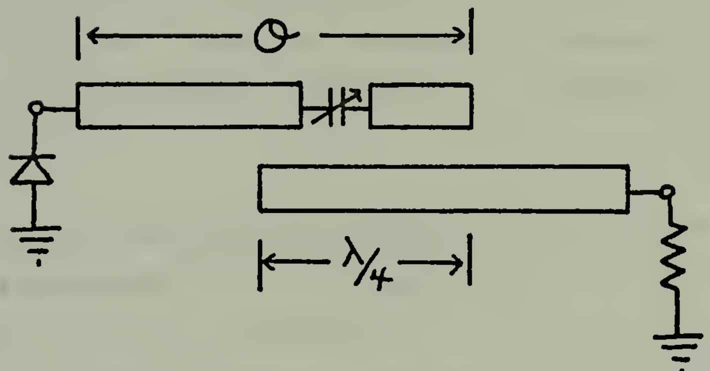
Varactor Model

Figure 18



Varactor in Parallel

Figure 19



Varactor in Series

Figure 20

H. PULSING CONSIDERATIONS

Techniques for pulsing the IMPATT oscillator were investigated, although no pulsing circuits were built nor were any of the stripline circuits previously described operated in anything other than the cw mode. Pulsing an IMPATT is a more difficult task than with a Gunn, requiring the pulsing circuit to be designed to mate specificaly with the particular IMPATT circuit[30].

There are basically two ways to pulse an IMPATT. Both methods require that bias be continuously applied to the diode just below the breakdown voltage (V_{BR} in table 3) and then a pulse voltage added to the bias voltage such that the IMPATT reaches the same value of operating current with each pulse. Since an IMPATT's power output and oscillator frequency change with bias current, the basic pulsing circuit would have to be a constant current generator. Although the thought of having to maintain continuous bias on the diode just below V_{BR} seems wasteful, it is an absolute must in order to keep from burning out the diode. The small pile of shorted diodes accumulated by this investigator verifies that the IMPATT is highly susceptible to transients.

One method of pulsing an IMPATT for durations of less than one microsecond is through a pulsing transformer as demonstrated by other investigators[6,31]. Peak powers much greater than cw specifications can be obtained and tolerated by the diode since the junction temperature is kept low. However, care must be taken not to overdrive the device or spectral distortion results.

The second method of pulsing involves the application of bias voltage through a transistor or SCR switching network[6]. Such circuits are necessary for clean spectra

at pulse widths exceeding one microsecond. The peak power will decrease with either increasing pulse width or duty cycle. Reference 6 indicates that double-drift IMPATTs are far superior in pulsed applications than are single-drift diodes. However, no information is available comparing pulsed double-drift silicon diodes to pulsed gallium arsenide IMPATTs.

IV. CONCLUSIONS AND RECOMMENDATIONS

A. CONCLUSIONS

The unsuccessful attempts at building a high power stripline or microstrip oscillator led to several hypotheses which remain unsubstantiated. The foundation for most of these is due to the fact that the diode package is considerably larger than that of other IMPATTs. When placed in a stripline or microstrip circuit, the packaging is physically larger than the strip. Hence, most of the rf power may be stored in the package reactance instead of being transferred to the load. The loss calculations seem to indicate that very little rf power is being dissipated in the dielectric surrounding the circuit. The large cathode cap may also be introducing a capacitance between it and the ground planes when inserted in the stripline circuit. The capacity could be so high when compared to the circuit's impedance that it provides an rf short to ground. The circuits built on the 3M board were smaller, thus making the previously described conditions more predominent. In contrast, coaxial circuits are built such that the cathode cap diameter is the same size as the circuit to which it connects. Also, the cathode cap and alumina encasement are not in proximity to nor surrounded by any grounded conductors as in the stripline circuits. Oddly enough, surrounding the diode with less ground plane material lowered the power output in the Rexolite circuits while removing the surrounding ground planes as in the 3M circuits resulted in no oscillations. A diode placed in a waveguide circuit is by comparison dwarfed by its surrounding circuitry. Industry has proven that solid state microwave devices in waveguide perform quite well.

The failure of the microstrip circuits is also

attributed to the size of the device packaging. One test of this premise would be to obtain the same diode from the manufacturer without the cathode cap and alumina encasement and put it in a microstrip configuration on a 10 to 25 mil thick dielectric slab. The heat sink stud would have to be retained to keep the diode junction temperature within specifications. The diode would be connected to the circuit by a gold bonding wire as it had been bonded to the cathode cap. The diode would be mounted immediately at the input strip to keep the wire inductance low. The mounting would be perpendicular (as in figure 7, but without the cap) such that the wire would be parallel to the ground plane. The facilities at NPGS and the experience of the author did not lend themselves to this investigation.

The circuits investigated were only to be used as starting points for the swept frequency oscillator. parallel coupled lines are not selective enough to eliminate the second and third harmonics from the IMPATT[21]. The addition of a non-linear device such as a varactor would certainly generate spurious products. Therefore, a lowpass or bandpass filter would be required in either microstrip or stripline circuits since the Q would be low. The investigation convinced the author that there is far more to designing microstrip or stripline circuitry employing active devices than meets the eye.

B. RECOMMENDATIONS

Investigation of stripline circuitry has led the author to believe that a waveguide or coaxial oscillator would prove more successful. The design of a varactor tuned coaxial oscillator may encounter some unexpected obstacles but Hewlett-Packard demonstrates that at least 1.25 watts output can be achieved from every 0425 IMPATT sold by testing each diode in a coaxial circuit. Strange as it may

seem, one source claims that coaxial circuits can be manufactured just as easily, and for less cost, than stripline or microstrip circuits[21]. Post-mounted diodes in waveguide have been demonstrated by other investigators, including some circuits with varactor tuners, whose work has been scrutinized and published by the _IEEE Transactions on Microwave Theory and Techniques_. The physical size of a waveguide or coaxial oscillator at X-band would not be so large as to make its positioning inside the body of the racon impossible. Regarding the specifications listed in table 2, frequency stability in stripline is harder to obtain compared to waveguide or coax circuits, and the temperature affects on a stripline circuit are more pronounced than the two alternatives[30]. The higher Q of the alternative circuits would also eliminate the need for exotic filtering.

Even though a one watt Gunn diode has yet to be announced, several smaller diodes can be combined in parallel to produce the desired output power. Such an arrangement would require the diodes to be matched so that they all turn on at the same time. Since multiple diodes increase expense, this disadvantage may be outweighed by the fact that if one diode ceases to function, the transmitter still operates, but at reduced power. A single source racon would be off the air completely should a diode failure occur.

One final alternative would be to use an updated version of the present transmitter design. A 20 watt, 2.3 GHz transistor could be used in a varactor multiplier to produce over 8 watts at 9.3 GHz. Assuming a transistor efficiency of 45% and a varactor conversion efficiency of 45%, the dc to rf efficiency would be about 20%. Or, a six watt, 3.1 GHz transistor could be used similarly. Either method may be able to take advantage of the existing power

supply, sweep, pulsing, and modulating circuits. The transmitter in the Model 227-X racon has already demonstrated the practicality of such an energy source.

In the opinion of the author, the varactor multiplier circuit may be the best approach from a cost-effective standpoint. If a state-of-the-art approach is desired, an IMPATT waveguide oscillator may eventually prove to be the best solution.

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